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<p>Efficient equivalent source representations were developed for coaxially shielded cables and for multiconductor cables. The correctness of these equivalent source representations was validated by linear and nonlinear terminations. The distribution of shield coupling parameters was studied for shielded multiconductor cables. The results of this study indicated that an approximate equivalent source excitation may be used for multiconductor cables, thereby reducing the source requirements even further. The approximate equivalent source excitation requires only two independent sources to represent the EMP response of the cable. The simplicity of the equivalent source excitation means that numerical models will run quicker, and system-level stimulation is possible with a lower power level and less complex simulators.</p>					
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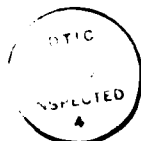
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1. INTRODUCTION

The electromagnetic pulse (EMP) analysis of a system frequently focuses on the response of cables excited by the transient fields. A large variety of analysis tools and techniques is available to aid in the study of the cable response. EMP simulators are available for experimental studies of system response. Difficulties arise, however, when we attempt to incorporate the results of cable coupling codes into circuit analysis codes for damage assessments. Similarly, obtaining threat-level responses with typical simulators is possible for only very short cables and compact systems. This report presents a solution to these problems using an equivalent source representation of the cable. Initially, the technique is developed for a single coaxially shielded cable. Then the approach is extended to multiconductor cables.

2. RESPONSE OF A COAXIALLY SHIELDED CABLE

The response of a coaxial cable due to a distributed excitation is reviewed in section 2.1. An equivalent and more compact source representation is developed in section 2.2. Section 2.3 investigates the validity of the equivalent representation when the line is loaded with nonlinear terminations.

2.1 Distributed Excitation

The response of a coaxially shielded cable with distributed excitation may be treated like the two-wire line problem shown in figure 1. The distributed voltage source, $V_s(x, \omega)$, is related to the complex transfer impedance of the shield and the external current flowing on it. The distributed current source, $I_s(x, \omega)$, accounts for coupling through the holes of the shield. Detailed explanations of the shield coupling mechanisms and solutions of the coupling problem are provided in numerous reports (e.g., [1], [2], [3]).* For this paper it is sufficient to define the general coupling problem as two arbitrary distributed sources--a series voltage source and a shunt current source. The transmission line of figure 1 has a characteristic impedance, K , a propagation characteristic, Γ , and linear terminations Z_1 and Z_2 at $x = 0, l$, respectively.

Schelkunoff [4] provides a solution for the terminal response of the transmission line shown in figure 1:

$$\begin{aligned} V(0) &= \int_0^l V_s(\xi) G_{V1}(0, \xi) d\xi + \int_0^l I_s(\xi) G_{V2}(0, \xi) d\xi \\ &= -\frac{KZ_1}{D} \int_0^l V_s(\xi) \{K \cosh \Gamma(l - \xi) + Z_2 \sinh \Gamma(l - \xi)\} d\xi - \end{aligned}$$

*References are listed at end of report.

$$\begin{aligned}
& - \frac{K^2 Z_1}{D} \int_0^1 I_s(\xi) \{K \sinh \Gamma(1 - \xi) + Z_2 \cosh \Gamma(1 - \xi)\} d\xi \\
V(1) &= \int_0^1 V_s(\xi) G_{V1}(1, \xi) d\xi + \int_0^1 I_s(\xi) G_{V2}(1, \xi) d\xi \\
&= \frac{K Z_2}{D} \int_0^1 V(\xi) \{K \cosh \Gamma \xi + Z_1 \sinh \Gamma \xi\} d\xi \\
& - \frac{K^2 Z_1}{D} \int_0^1 I_s(\xi) \{K \sinh \Gamma \xi + Z_1 \cosh \Gamma \xi\} d\xi
\end{aligned}$$

where

$$D = K \left[(K^2 + Z_1 Z_2) \sinh \Gamma l + K(Z_1 + Z_2) \cosh \Gamma l \right] \quad (1)$$

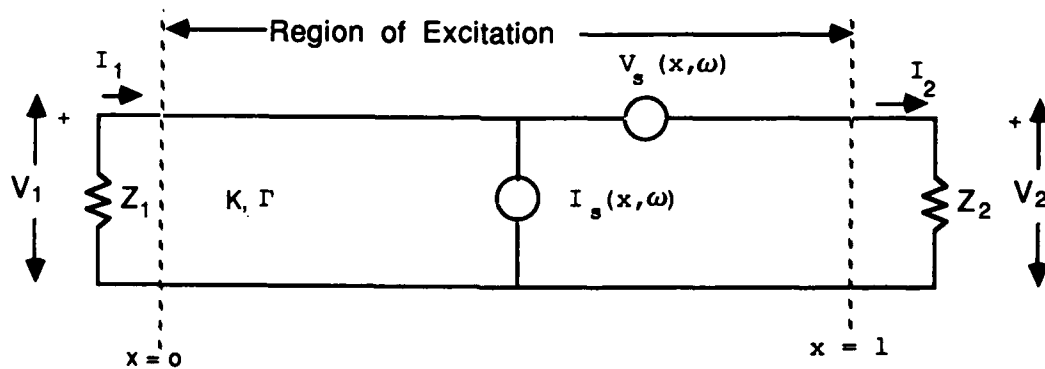


Figure 1. Transmission line with distributed excitation.

The line response may be computed using the proper frequency descriptions of the line parameters, terminations, and source functions. The time domain response is then obtained using a suitable inverse Fourier transform.

2.2 Equivalent Excitation

It is desired to develop an equivalent model of the transmission line problem of figure 1. Characteristics of the model should include a simplified source characterization, ease of implementation in circuit analysis codes, applicability to nonlinear loading, and a relatively easy physical implementation for testing purposes. Equivalent source theory will be reviewed briefly before an alternate coaxial cable coupling model is developed.

2.2.1 Thevenin's and Norton's Theorems

Both Thevenin's and Norton's theorems deal with the development of equivalent source representations of a linear network, N , that excites an arbitrary load network, N' , as shown in figure 2. Thevenin's theorem states that the network N may be replaced by an equivalent generator consisting of a voltage source and a series internal impedance such that the terminal response (current and voltage at T_1, T_2) will be identical to the original network (N and N'). An equivalent current source representation may be developed through the application of Norton's theorem.

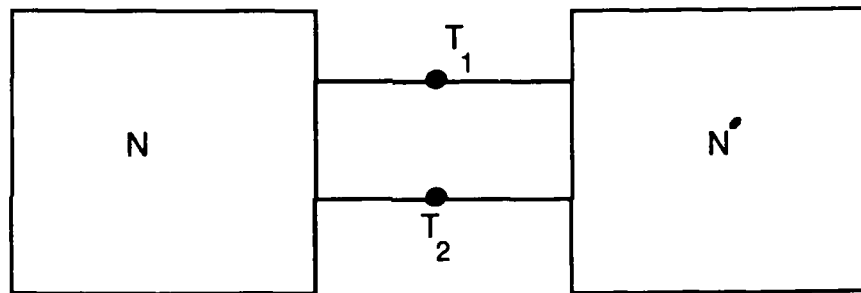


Figure 2. Interconnected networks.

Peskin [5] develops Thevenin's and Norton's theorems subject to the following assumptions:

- (a) The network N in figure 2 is linear.
- (b) The network N is stable over the full range of load values possible.
- (c) The terminal response is Laplace transformable; this does not require N' to be linear; only its load voltage and current must obey

$$\int_0^{\infty} |f(t)| e^{-\sigma t} dt < \infty ,$$

for some real number, σ .

Using these assumptions and the node relationships, it is a relatively straightforward matter to prove that the network N may be replaced by a voltage source equal to the open circuit voltage at terminals T1,T2 and a series impedance equal to the input impedance of N looking into T1,T2. The input impedance is found by shorting out all independent voltage sources and opening all independent current sources. The Norton equivalent uses a current generator equal to the short circuit current at T1,T2 shunted by the input admittance (the reciprocal of the input impedance).

This two-terminal or single port theorem can be extended to multi-port networks if we require m independent sources, one for each of m ports, and if we define an impedance (admittance) matrix composed of the self impedance (admittance) and transfer impedance (admittance) from each of the other ports. The source voltages for a Thevenin equivalent network are defined by

$$V_k^{oc} = V_k, \quad I_j = 0, \quad j=1, m, \quad (2)$$

where V_k is the potential at port k and I_j is the current through port j. Similarly, the Norton current sources are the short circuit currents at each port when all the ports are shorted. The impedance matrix for the Thevenin representation is defined by the equation

$$z_{ij} = V_i / I_j, \quad I_k = 0, \quad \forall K \neq j, \quad (3)$$

under the condition that all internal independent sources are properly shut off. The admittance matrix is the inverse of the impedance matrix. The multiport Thevenin theorem will be applied in the next section to develop an equivalent representation of the coaxially shielded cable.

2.2.2 Thevenin Equivalent Representation

Modeling the coaxial cable response with a Thevenin equivalent generator requires computation of the open circuit voltage at each end. Using equation (1) and letting Z_1, Z_2 approach infinity yields the following open circuit voltages:

$$V_{oc}(0) = - \int_0^1 V_s(\xi) \frac{\sinh \Gamma(1-\xi)}{\sinh \Gamma} d\xi - K \int_0^1 I_s(\xi) \frac{\cosh \Gamma(1-\xi)}{\sinh \Gamma} d\xi$$

and

$$V_{oc}(1) = \int_0^1 V_s(\xi) \frac{\sinh \Gamma \xi}{\sinh \Gamma} d\xi - K \int_0^1 I_s(\xi) \frac{\cosh \Gamma \xi}{\sinh \Gamma} d\xi. \quad (4)$$

Figure 3 shows the equivalent network. The impedance matrix, Z , is simply

$$\bar{Z} = \begin{bmatrix} K/\tanh \Gamma & -K/\sinh \Gamma \\ K/\sinh \Gamma & -K/\tanh \Gamma \end{bmatrix}, \quad (5)$$

and the terminal response is given by

$$\bar{V} = \begin{bmatrix} V_1 \\ V_2 \end{bmatrix} = \bar{Z} \begin{bmatrix} I_1 \\ I_2 \end{bmatrix} + \begin{bmatrix} V_s^1 \\ V_s^2 \end{bmatrix} = \bar{Z}\bar{I} + \bar{V}_s. \quad (6)$$

Using the impedance relationship for the load voltage and current, $I_1 = -V_1/Z_1$ and $I_2 = V_2/Z_2$, the load voltage is found from equation (6):

$$\bar{V} = \bar{Z} \begin{bmatrix} -1/z_1 & 0 \\ 0 & 1/z_2 \end{bmatrix} \bar{V} + \bar{V}_s. \quad (7)$$

Equation (7) may be solved for V as follows:

$$\begin{aligned} V &= \left[\bar{u} - \bar{Z} \begin{bmatrix} -1/z_1 & 0 \\ 0 & 1/z_2 \end{bmatrix} \right]^{-1} \bar{V}_s \\ &= \begin{bmatrix} 1 + K/(z_1 \tanh \Gamma) & K/(z_2 \sinh \Gamma) \\ K/(z_1 \sinh \Gamma) & 1 + K/(z_2 \tanh \Gamma) \end{bmatrix}^{-1} \bar{V}_s \\ &= \frac{K}{D} \begin{bmatrix} z_1 [z_2 \sinh \Gamma + K \cosh \Gamma] & -z_1 \\ -z_2 & z_2 [z_1 \sinh \Gamma + K \cosh \Gamma] \end{bmatrix} \bar{V}_s, \quad (8) \end{aligned}$$

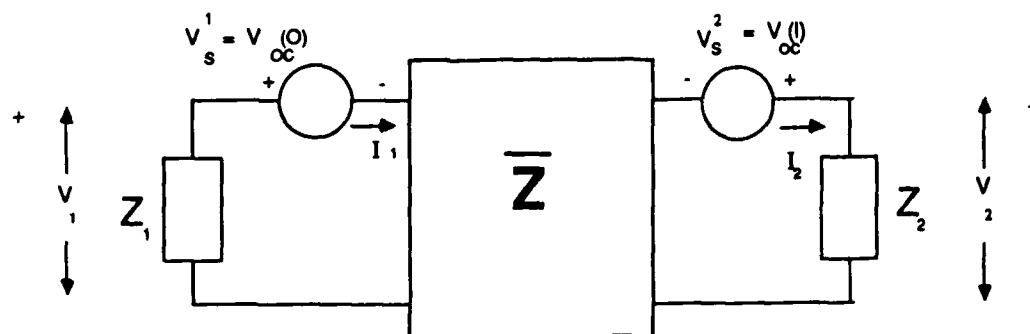


Figure 3. Equivalent network for coaxial cable.

where \bar{u} is the identity matrix. After substitution of equation (4) for the source term, \bar{V}_s , manipulation of equation (8) yields equation (1), demonstrating the correctness of the equivalent network.

The advantages of the equivalent representation are that the distributed source is replaced by two point sources and the excitation is separated from the transmission line. This means that standard transmission-line models available in most circuit analysis codes may be used in conjunction with two voltage sources to model the coaxial cable. Similarly, from a simulation standpoint, an injection system may be developed using two independent voltage sources, one for each end of the cable. Therefore, the objective of developing a simplified representation for the EMP coupling to a coaxial cable has been achieved. The one area not fully explored is the correctness of the equivalent representation when nonlinear loads appear at the terminals in place of Z_1 and Z_2 . Nonlinear loading is studied in the next section.

2.2.3 Nonlinear Terminations

It was demonstrated that the equivalent Thevenin representation for the coaxial cable produced results identical to the distributed excitation when linear terminations are used. While the single port representation is valid for nonlinear loads, nothing has been done to validate the multi-port representation of distributed networks with nonlinear loads. In the following analysis it is assumed that the transmission line is linear and that the potentials created because of the EMP excitation do not exceed the dielectric breakdown voltages for the cable. Note that this does not preclude the inclusion of nonlinear effects on the exterior of the cable; only its internal line must be linear. If internal breakdown is possible, then the model would have to be further partitioned to isolate the nonlinear regions.

Obtaining a closed form solution for the response of a transmission line with nonlinear loads is in all probability impossible. However, if a slight restriction is placed on the nature of the terminations, the problem becomes fairly straightforward, and useful insight may be gained into the actual line response. As illustrated in figure 4, it is possible to very closely approximate a nonlinear load characteristic with a set of linear curves. The piecewise-linear approach will be used to study the validity of the equivalent transmission line representation under nonlinear loading.

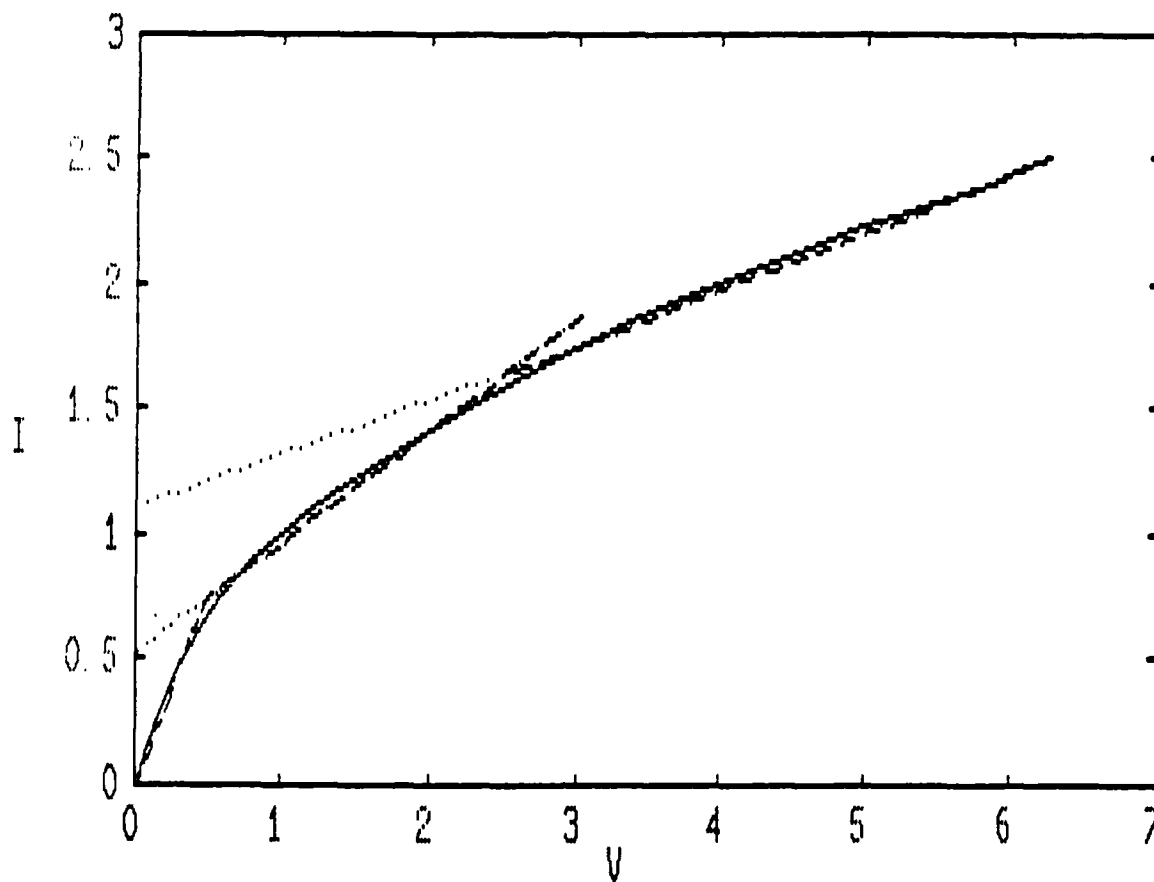


Figure 4. Piecewise-linear representation of nonlinear load.

Solving for the response with nonlinear loads simply involves a series of linear solutions, with the loading over a specific time period being determined by the load voltage or current. The only complication in the solution is the inclusion of initial condition effects at transition points.

An example of how a piecewise-linear solution is used is presented in figure 5. Here the response of a circuit loaded by the simple square law device of figure 4 is computed. As the load current, I , reaches a transition value of the piecewise-linear model, a new linear equation must be solved. In practice, only a finite number of segments will be used to represent the nonlinear load. However, if computation load is not a consideration, then a very large set of linear curves may be used to achieve any desired accuracy. The current response shown in figure 5 was calculated with only two linear regions. The piecewise-linear response compared within one percent with the response computed with the MICROCAP-II circuit analysis code.

The solution for the EMP response of the coaxial cable given in equation (1) assumed that the initial conditions were zero. The effect of the nonzero initial conditions at different points in time is seen in the transforms of the differential equations describing the transmission line

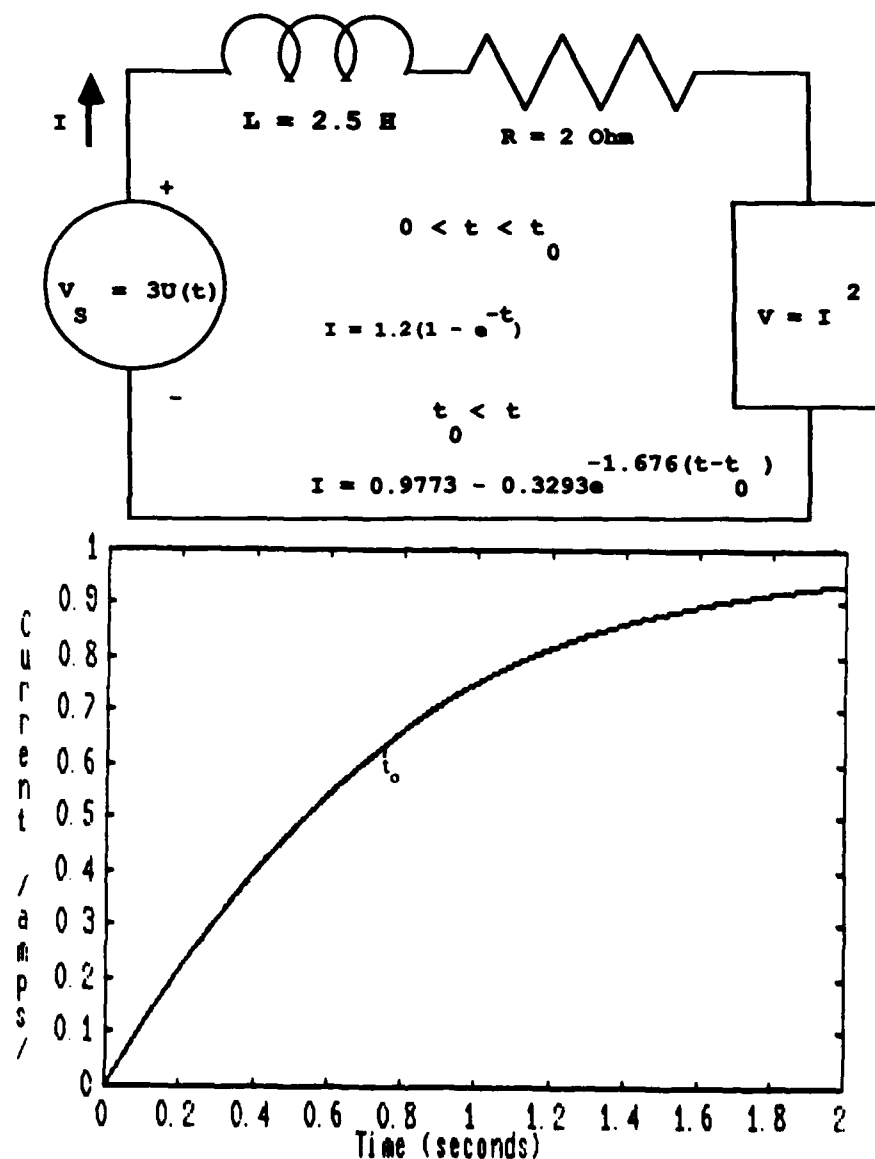


Figure 5. Piecewise-linear solution.

$$\frac{dV}{dx} = -(R + j\omega L)I + LI(t_0) + V_s$$

$$\frac{dI}{dx} = -(G + j\omega C)V + CV(t_0) + I_s. \quad (9)$$

The effect of the initial conditions is to add constant source terms, $LI(t_0)$ and $CV(t_0)$, to the EMP sources, V_s and I_s . Defining two new source terms,

$$V'_s = V_s + L I(t_0)$$

and

$$I'_s = I_s + C V(t_0); \quad (10)$$

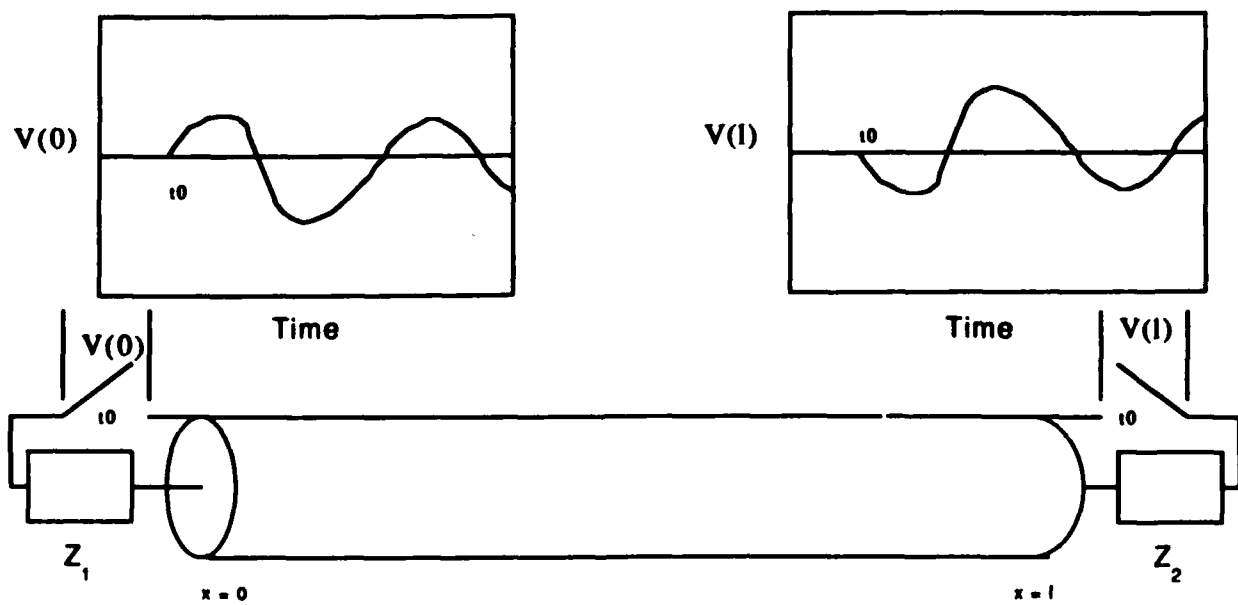
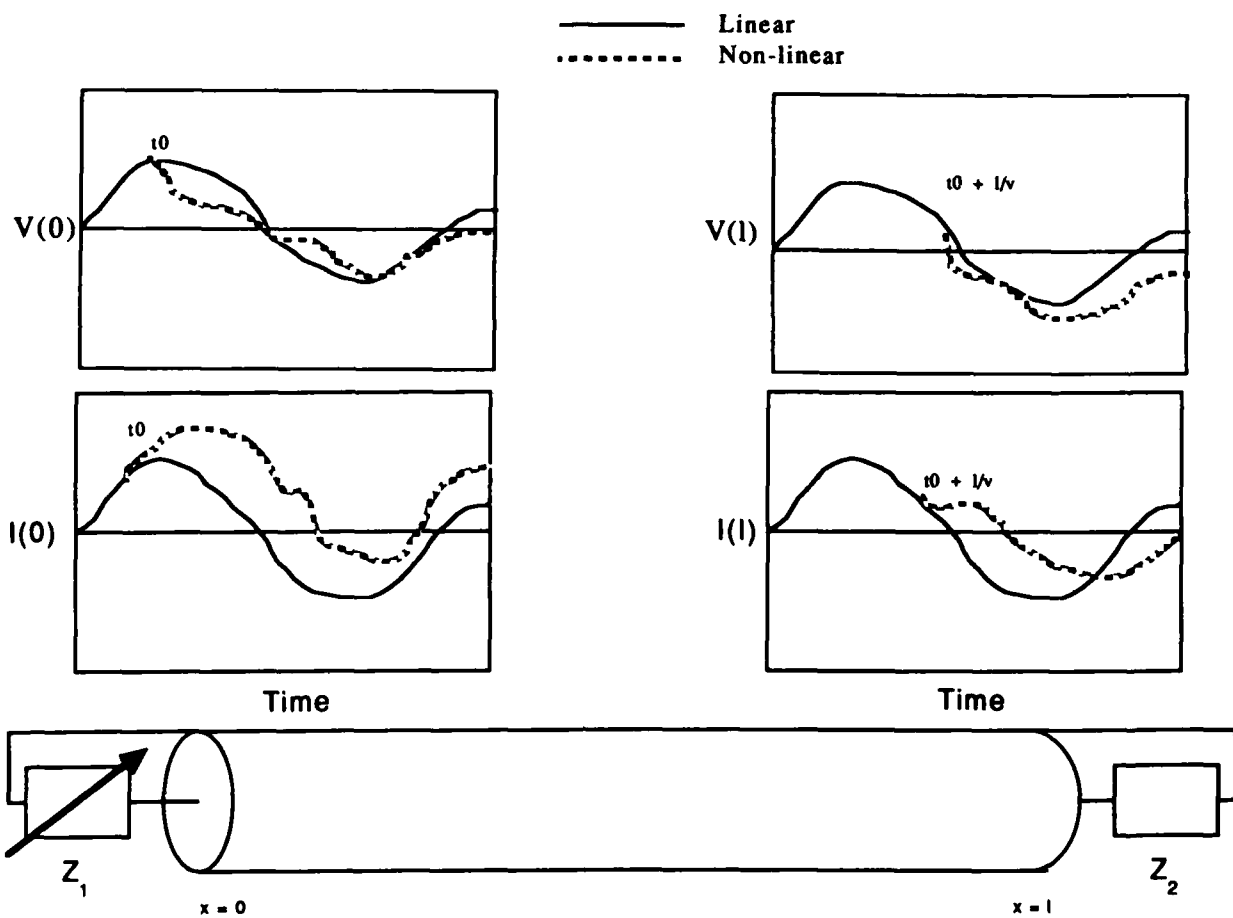
the transmission-line response is given by equation (1) with the new source terms. Therefore, from a computational standpoint, a new Thevenin equivalent generator is needed to model the line after a transition has occurred in one of the terminations. The impedance matrix, Z , remains unchanged, but the sources at each end are now

$$\begin{aligned} V_{oc}(0) &= - \int_0^l V'_s(\xi) \frac{\sinh \Gamma(l - \xi)}{\sinh \Gamma l} d\xi - K \int_0^l I'_s(\xi) \frac{\cosh \Gamma(l - \xi)}{\sinh \Gamma l} d\xi \\ V_{oc}(l) &= \int_0^l V'_s(\xi) \frac{\sinh \Gamma \xi}{\sinh \Gamma l} d\xi - K \int_0^l I'_s(\xi) \frac{\cosh \Gamma \xi}{\sinh \Gamma l} d\xi \end{aligned} \quad (11)$$

Therefore, the criterion for determining the validity of the Thevenin equivalent network with nonlinear loading is whether or not it produces the open circuit voltages defined in equation (11) after load transitions. The question arises not because of the distributed EMP sources, V_s and I_s , but rather because of the initial conditions, $I(t_0)$ and $V(t_0)$. Obviously, the transmission-line current and voltage distributions are different for the actual EMP excitation and its Thevenin equivalent. If the proper open circuit voltage is produced after time t_0 , then the Thevenin equivalent network is valid even when nonlinear loads are used.

One method of verifying the correctness of the open circuit voltages would be to (1) compute the complete line response under both types of excitation, (2) transform the response into the time domain to obtain the initial condition distribution along the line, and (3) solve for the open circuit voltages with the distributed initial conditions, $V(t_0)$ and $I(t_0)$, and EMP source terms, V_s and I_s . Unfortunately, it is very difficult to verify open circuit voltages, even for a simple circuit. A similar approach was used to analyze a two-section lumped parameter transmission-line model [6]. Based on the complexity encountered with only a simple circuit, it is not likely that a general solution can be obtained. Therefore, a different approach is needed.

Assume that the terminal responses of a coaxial cable are as shown in figure 6. Let there be a nonlinearity in the termination at $x = 0$ such that the impedance, Z_1 , changes due to the voltage at time t_0 . Next, assume that the open circuit voltages from time t_0 on are as shown in figure 7. Note that these open circuit voltages are different from the voltages starting at time t_0 due to the loading from time zero to t_0 . These open circuit voltages may now be used to compute the response of the line after the nonlinear transition at time t_0 as shown in figure 6 by the dashed waveforms.



Using contradiction, we will now prove the validity of using a Thevenin equivalent network to model the response of a transmission line with nonlinear loading. In section 2.2.2 it was shown that the response of the Thevenin network would duplicate the line response with linear loads (the solid curves in fig. 6). Now assume that the open circuit voltages of the equivalent circuit after time t_0 are not the same as figure 7. This is the only way the equivalent network will not provide the correct response after the transition at time t_0 . We now have two different networks for times greater than t_0 , one with the correct open circuit voltages from the actual transmission line and one with incorrect open circuit voltages from the original Thevenin equivalent network. The impedance networks are identical because of the assumption that the line is linear and operating below breakdown. In section 2.2.2 it was shown that a Thevenin equivalent network will provide correct terminal responses for any linear loading. Now reconnect the original terminations, Z_1 and Z_2 , to the new equivalent network. Since it was assumed that the open circuit voltages from the original Thevenin model are incorrect, the response with the original loading is also incorrect. This violates the work of section 2.2.2 that proved that the Thevenin equivalent would provide the proper response for all time with a linear load. Therefore, the assumption must be false that the open circuit voltages produced by the equivalent network are different from the open circuit voltages produced by the actual distributed excitation.

An alternate approach for validating the Thevenin equivalent network can be developed that uses the uniqueness of the solution for the transmission line differential equations.

Since the differential equations governing the transmission line are linear and time invariant, the solution for a given input and initial state is unique. This implies that the input required to produce a specified output from an initial condition set is also unique.

The open circuit voltages, V_s' , required to produce a given terminal response, $V_1(t_0)$ and $V_2(t_0)$, starting from time, t_0 , are found from equation (8) to be

$$\bar{V}_s' = \begin{bmatrix} V_s^1(t_0) \\ V_s^2(t_0) \end{bmatrix} = \begin{bmatrix} \bar{u} - \bar{Z} \begin{bmatrix} -1/z_1 & 0 \\ 0 & 1/z_2 \end{bmatrix} \end{bmatrix} \begin{bmatrix} V_1(t_0) \\ V_2(t_0) \end{bmatrix}, \quad (12)$$

where the voltages $V_1(t_0)$ are the Fourier transforms of the time domain waveforms from time t_0 on. Since the actual network response and its Thevenin equivalent produce identical terminal responses with linear loads, Z_1 and Z_2 , the open circuit voltages from time t_0 must be identical. Otherwise, more than one set of inputs will provide the same outputs. This is not possible with a linear line. The effective open circuit voltages starting at an arbitrary time t_0 due to the impressed sources and the initial conditions must be the same for the actual line with distributed excitation and its Thevenin counterpart with only two source terms. Therefore, both lines will provide the same terminal response after a termination transition at a time, t_0 .

Equation (12) suggests a method for solving for the response of a coaxial line with nonlinear terminations using Fourier transform techniques. The load nonlinearity is assumed to be piecewise-linear. The steps to solve for the nonlinear response are as follows:

- (a) Compute the complete terminal response with the nominal values for the terminations. This is done in the frequency domain and transformed into the time domain.
- (b) Determine the time, t_0 , where a transition will occur in a termination.
- (c) Compute the effective open circuit voltage with a start time of t_0 . These voltages are found by computing the Fourier transforms of the terminal responses multiplied by a unit step function starting at t_0 . Alternatively this may be accomplished by convolution of the frequency response at the terminals with a unit step defined by

$$\bar{V}'_s = \left[\bar{u} - \bar{z} \begin{bmatrix} -1/z_1 & 0 \\ 0 & 1/z_2 \end{bmatrix} \right] \left[\bar{f} \left\{ u(t-t_0) * \begin{bmatrix} V_1 \\ V_2 \end{bmatrix} \right\} \right] \quad (13)$$

- (d) Solve for the terminal response after the transition ($t \geq t_0$) using equation (8), the source voltages from step 3, and the new terminations.

This is a simple approach to solving a complex problem. This technique should be computationally superior to finite difference techniques when the line lengths are long and many cells would be needed to model the transmission line.

3. RESPONSE OF A MULTICONDUCTOR TRANSMISSION LINE

The equivalent source representation developed in section 2 for simple coaxial cables will now be extended to multiconductor transmission lines. After a brief review of multiconductor transmission line theory in section 3.1, the equivalent source representation is developed in section 3.2. Several implementation techniques are reviewed in section 3.3.

3.1 Distributed Excitation

Consider the lossless multiconductor transmission line shown in figure 8. The line consists of $N + 1$ conductors, with conductor $N + 1$ being the reference conductor. The length of the line is l , and its electrical

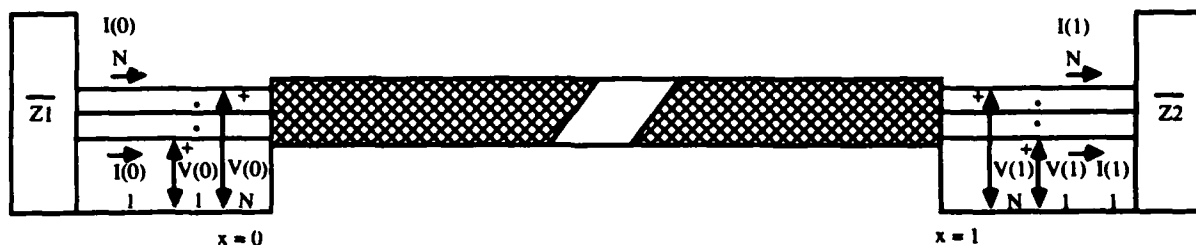


Figure 8. Multiconductor transmission line.

characteristics are determined by the capacitive coefficient matrix C and the inductive coefficient matrix L . The line is described by the following matrix partial differential equations [7]:

$$\begin{aligned} \frac{\partial \bar{v}(x,t)}{\partial x} &= -\bar{L} \frac{\partial \bar{i}(x,t)}{\partial t} \\ \frac{\partial \bar{i}(x,t)}{\partial x} &= -\bar{C} \frac{\partial \bar{v}(x,t)}{\partial t} \end{aligned} \quad (14)$$

when no source terms are present. Using the Laplace transform these equations become:

$$\frac{\partial}{\partial x} \begin{bmatrix} \bar{V} \\ \bar{I} \end{bmatrix} = -s \begin{bmatrix} \bar{0} & \bar{L} \\ \bar{C} & \bar{0} \end{bmatrix} \begin{bmatrix} \bar{V} \\ \bar{I} \end{bmatrix}, \quad (15)$$

where the dependence of \bar{V} and \bar{I} on x is implicit; \bar{V} and \bar{I} are $n \times 1$ vectors; \bar{L} and \bar{C} are $n \times n$ matrices and $\bar{0}$ is a zero matrix of order $n \times n$.

Differentiating equation (14) and substituting yield a set of decoupled second-order differential equations:

$$\frac{\partial^2}{\partial x^2} \begin{bmatrix} \bar{V} \\ \bar{I} \end{bmatrix} = +s^2 \begin{bmatrix} \bar{L}\bar{C} & \bar{0} \\ \bar{0} & \bar{C}\bar{L} \end{bmatrix} \begin{bmatrix} \bar{V} \\ \bar{I} \end{bmatrix}. \quad (16)$$

It is possible for lossless multiconductor cables to define a transformation matrix T such that

$$s^2 \bar{T}^{-1} \bar{C} \bar{L} \bar{T} = \gamma^2, \quad (17)$$

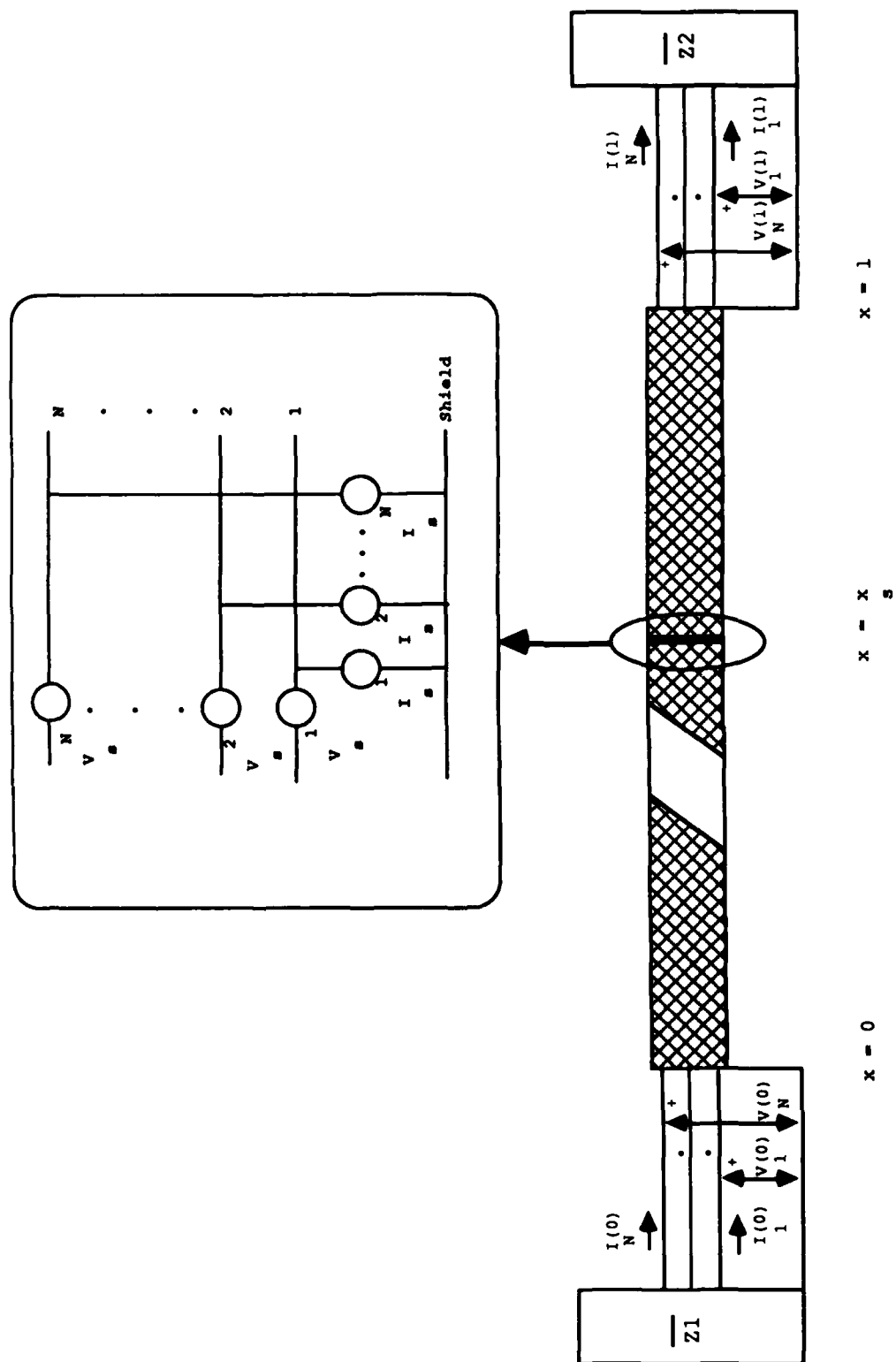


Figure 9. Generalized point source for multiconductor transmission line.

where $\bar{\gamma}$ is an $n \times n$ diagonal matrix with elements

$$\gamma_{ij}^2 = \begin{cases} \gamma_i^2, i = j \\ 0, i \neq j \end{cases}, \quad (18)$$

and γ_1 an eigenvalue of equation (16).

Then using the transformations $\bar{V} = \bar{T}\bar{V}'$ and $\bar{I} = \bar{T}\bar{I}'$, equation (16) becomes a set of $2N$ decoupled differential equations with a solution given by

$$\begin{bmatrix} \bar{V} \\ \bar{I} \end{bmatrix} = \begin{bmatrix} \bar{Z}_c & -\bar{Z}_c \\ \bar{U} & \bar{U} \end{bmatrix} : \begin{bmatrix} \bar{T}e^{-\bar{\gamma}x} \bar{\alpha}^+ \\ \bar{T}e^{-\bar{\gamma}x} \bar{\alpha}^- \end{bmatrix}, \quad (19)$$

where

$$\bar{Z}_c = s^{-1} \bar{C}^{-1} \bar{T} \bar{\gamma} \bar{T}^{-1} \quad (20)$$

is the characteristic impedance matrix, \bar{U} is the identity matrix, and $\bar{\alpha}^+$ and $\bar{\alpha}^-$ are $n \times 1$ vectors determined by the terminating impedances and the source terms exciting the line. Tesche et al [7] provide a solution for $\bar{\alpha}^+$ and $\bar{\alpha}^-$ when the line is terminated in impedance matrices Z_1, Z_2 at $x = 0$ and $x = 1$, respectively, and in a generalized point source at x_s , as shown in figure 9. Of interest here is the terminal response of the line. The voltage and current at the ends are related by the terminating matrices as follows:

$$\begin{bmatrix} \bar{V}(0) \\ \bar{V}(1) \end{bmatrix} = \begin{bmatrix} -\bar{Z}_1 & \bar{0} \\ \bar{0} & \bar{Z}_2 \end{bmatrix} : \begin{bmatrix} \bar{I}(0) \\ \bar{I}(1) \end{bmatrix}, \quad (21)$$

and from Frankel [8],

$$\begin{bmatrix} \bar{I}(0) \\ \bar{I}(1) \end{bmatrix} = \begin{bmatrix} \bar{U} - \bar{\Gamma}_1 & \bar{0} \\ \bar{0} & \bar{U} - \bar{\Gamma}_2 \end{bmatrix} : \begin{bmatrix} \bar{\Gamma}_1^{-1} & \bar{T}e^{\bar{\gamma}} \bar{T}^{-1} \\ \bar{T}e^{\bar{\gamma}} \bar{T}^{-1} & \bar{\Gamma}_2^{-1} \end{bmatrix} : \begin{bmatrix} \bar{\delta}^- \\ \bar{\delta}^+ \end{bmatrix}, \quad (22)$$

where

$$\bar{\Gamma}_i = [\bar{Z}_i + \bar{Z}_c]^{-1} [\bar{Z}_i - \bar{Z}_c] \quad (23)$$

is the reflection coefficient associated with load impedance Z_1 , and $\bar{\delta}^-$ and $\bar{\delta}^+$ are source-related terms. The operator ":" in the above equations arises from the fact that the elements of the matrices are also matrices. The matrices are first multiplied as if they contained normal scalar elements. Then the resulting matrix operations for each element are carried out.

The source terms $\bar{\delta}^-$ and $\bar{\delta}^+$ have been presented [9] for a generalized point source (fig. 9) as

$$\begin{bmatrix} \bar{\delta}^- \\ \bar{\delta}^+ \end{bmatrix} = 1/2 \begin{bmatrix} \bar{T} e^{\bar{\gamma} x_s} \bar{T}^{-1} \end{bmatrix} : \begin{bmatrix} \bar{Z}_c^{-1} & \bar{U} \\ \bar{Z}_c^{-1} & -\bar{U} \end{bmatrix} : \begin{bmatrix} \bar{V}_s \\ \bar{I}_s \end{bmatrix}, \quad (24)$$

where x_s is the location of the source. If a distributed source excites the line then equation (22) must be integrated over the source:

$$\bar{\delta}^\pm = \int_0^1 \bar{\delta}^\pm(\xi) d\xi, \quad (25)$$

and the voltage and current sources (V_s and I_s) are per unit length quantities.

3.2 Equivalent Excitation

As with the single coaxial cable, it is desired to develop an equivalent representation that readily lends itself to circuit analysis and point source simulation. The Thevenin equivalent network for the multiconductor cable is a straightforward extension of the coaxial cable development. The source at each end must be an n -dimensional vector to satisfy the most general conditions. The $2n$ port impedance matrix now requires $2n \times 2n$ terms to account for the self and transfer impedances of all the ports. The correctness of the multiconductor Thevenin equivalent network will be examined first. Then the impact of a restriction on the Thevenin source terms will be investigated. The purpose of the source restriction is to further simplify the model.

The response of the shielded multiconductor cable to the Thevenin equivalent source excitation may be obtained from equations (21) through (25). The layout of the Thevenin equivalent network is given in figure 10, where $V_s^1(0)$ and $V_s^2(1)$ are the wire to reference open-circuit voltage vectors at $x = 0$, $x = 1$, due to the distributed EMP excitation:

$$\begin{bmatrix} \bar{\delta}_d^- \\ \bar{\delta}_d^+ \end{bmatrix} = \frac{1}{2} \int_0^1 \begin{bmatrix} \bar{T} e^{\bar{\gamma} \xi} \bar{T}^{-1} \end{bmatrix} : \begin{bmatrix} \bar{Z}_c^{-1} & \bar{U} \\ \bar{Z}_c^{-1} & -\bar{U} \end{bmatrix} : \begin{bmatrix} \bar{V}_s \\ \bar{I}_s \end{bmatrix}. \quad (26)$$

3.2.1 Thevenin Source Terms

The open circuit voltages $V^{OC}(0)$ and $V^{OC}(1)$ are found from equations (21) through (23) and equation (26) if we let the terminating impedances between each wire and the reference conductor, as well as the wire-to-wire impedances, approach infinity (open circuit). The impedance matrix Z_1 may be found from the definition

$$Z_{ij} = V_i/I_j, I_k = 0 \forall k \neq j, \quad (27)$$

yielding

$$Z_{ij} = \begin{cases} \infty, i=j \\ 0, i \neq j \end{cases} \quad (28)$$

The source vector $[\bar{\delta}^-, \bar{\delta}^+]^T$ in equation (26) is independent of the terminating matrices.

The elements of equations (21) through (23) that depend on the termination matrices are of the form

$$\bar{Z}_i [\bar{U} - \bar{\Gamma}_i] = 2\bar{Z}_i [\bar{Z}_i + \bar{Z}_c]^{-1} \bar{Z}_c \quad (29)$$

and

$$\bar{\Gamma}_i = [\bar{Z}_i + \bar{Z}_c]^{-1} [\bar{Z}_i - \bar{Z}_c] \quad (30)$$

In the limit (\bar{Z}_i defined by eq (28)), it is found that equation (29) approaches $2\bar{Z}_c$. The reflection matrix, $\bar{\Gamma}_i$, approaches the unit matrix, \bar{U} , under open circuit conditions.

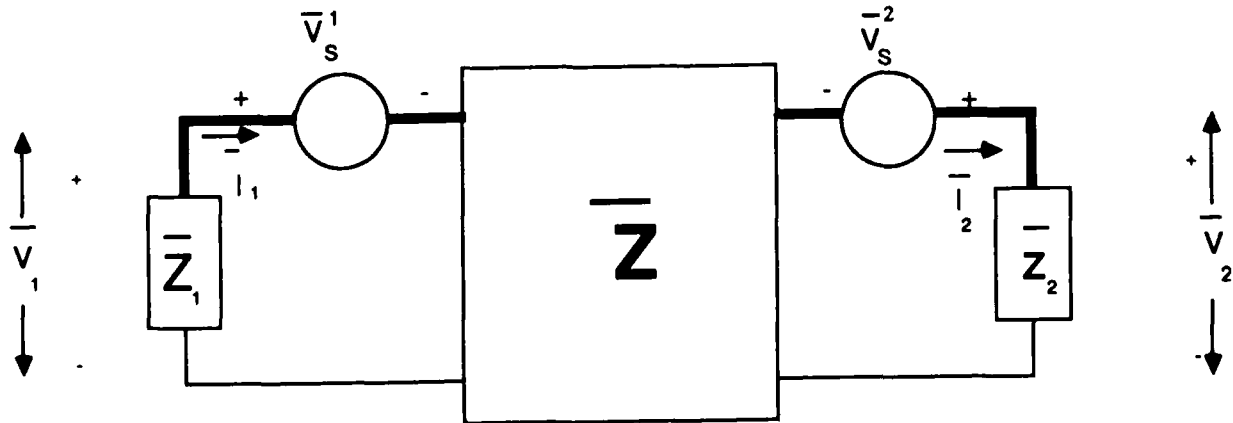


Figure 10. Equivalent source representation for multiconductor transmission line.

Using equations (21) and (22) to define the terminal voltages and taking the limit as the terminating impedances approach infinity yields the following open circuit voltages at the terminals:

$$\begin{bmatrix} \bar{V}^{\infty}(0) \\ \bar{V}^{\infty}(l) \end{bmatrix} = \begin{bmatrix} -2\bar{z}_c & \bar{o} \\ \bar{o} & 2\bar{z}_c \end{bmatrix} : \begin{bmatrix} \bar{U} & \bar{T}e^{\bar{\gamma}}\bar{T}^{-1} \\ \bar{T}e^{-\bar{\gamma}}\bar{T}^{-1} & \bar{U} \end{bmatrix}^{-1} : \begin{bmatrix} \bar{\delta}_d^- \\ \bar{\delta}_d^+ \end{bmatrix}, \quad (31)$$

with the source terms $\bar{\delta}_d^-$ and $\bar{\delta}_d^+$ as given in equation (26).

3.2.2 Terminal Response

The terminal response for the Thevenin equivalent network is found by superposition of the response from the source at each end of the cable. The source voltages given in equation (31) are used in conjunction with equation (24) to define the source terms $\bar{\delta}_p^-$ and $\bar{\delta}_p^+$ for the current responses given in equation (22). The current source, I_s , in equation (24) is zero because only a voltage source vector is used at each end of the cable. Summing the source terms due to the voltage sources at $x_s = 0, l$ yields

$$\begin{aligned} \begin{bmatrix} \bar{\delta}_p^- \\ \bar{\delta}_p^+ \end{bmatrix} &= 1/2 \begin{bmatrix} \bar{z}^{-1} & \bar{T}e^{\bar{\gamma}}\bar{T}^{-1}\bar{z}_c^{-1} \\ -\bar{T}e^{-\bar{\gamma}}\bar{T}^{-1}\bar{z}_c^{-1} & \bar{z}_c^{-1} \end{bmatrix} : \begin{bmatrix} \bar{V}^{\infty}(0) \\ \bar{V}^{\infty}(l) \end{bmatrix} = \\ &= \begin{bmatrix} \bar{U} & \bar{T}e^{\bar{\gamma}}\bar{T}^{-1} \\ \bar{T}e^{-\bar{\gamma}}\bar{T}^{-1} & \bar{U} \end{bmatrix} : \begin{bmatrix} \bar{U} & \bar{T}e^{\bar{\gamma}}\bar{T}^{-1} \\ \bar{T}e^{-\bar{\gamma}}\bar{T}^{-1} & \bar{U} \end{bmatrix}^{-1} : \begin{bmatrix} \bar{\delta}_d^- \\ \bar{\delta}_d^+ \end{bmatrix} \\ &= \begin{bmatrix} \bar{\delta}_d^- \\ \bar{\delta}_d^+ \end{bmatrix}, \end{aligned} \quad (32)$$

where $\begin{bmatrix} \bar{\delta}_d^- & \bar{\delta}_d^+ \end{bmatrix}^T$ is the distributed source term vector. This shows that the Thevenin equivalent network provides the same excitation at the terminals as the distributed excitation. The terminal response is given by equations (21) and (22).

3.2.3 Nonlinear Terminations

The nonlinear termination study for a single coaxial line (sect. 2.2.3) is directly applicable to multiconductor lines. The only real question was whether or not Thevenin equivalent circuits of transmission lines are valid with nonlinear terminations. A multiconductor cable simply requires a more complex source and impedance representation. Therefore, the work of section 2.2.3 implies that equivalent circuit representations are valid for multiconductor transmission lines with nonlinear terminations.

3.2.4 Coupling Distribution

In the previous sections it was shown that an equivalent source network of $2N$ independent voltage terms will provide the same terminal response as the distributed excitation. The distribution of the shield coupling parameters among the N conductors will now be examined in an attempt to further reduce the source requirements.

The distribution of the coupling through the shield will determine the source terms V_s and I_s in equation (22). Unfortunately, little work exists on the nature of the coupling distributions. As pointed out in reference [8], most shield coupling studies, both experimental and theoretical, consider coaxially shielded cables with only one conductor. Frankel [8] suggests measurement techniques, but does not provide any insight into the distribution.

In Tigner et al [9], a theoretical treatment is proposed for the distributions of the capacitive and inductive coupling parameters; the common mode inductive and capacitive sources are distributed over the conductors using coupling ratio vectors c and l such that

$$\sum_{i=1}^N C_i = 1, \quad \sum_{i=1}^N l_i = 1. \quad (33)$$

The requirement of equation (33) appears plausible for the capacitive coupling parameter since in effect the measured common mode current source is really a set of N current sources in parallel, each driving one of the wires. However, the requirement that the N multiconductor voltage sources add up to the common mode voltage sources appears to be inconsistent since the N voltage sources are also in parallel, not in series as equation (33) suggests. This appears on average to divide the voltage source amplitude by the number of conductors. Evidence of this improper scaling is seen in the results presented by Paul [10] for a comparison of the theoretical model with experimental data. A very large shield transfer impedance of 1 ohm/meter was derived empirically for a 37-conductor cable with a tape-wrapped shield. It is felt that this large shield transfer impedance is an artifact of the requirement of equation (33).

Unfortunately, no other models or experimental data have been discovered in the literature. The following model development attempts to overcome the difficulties discussed above. Only the inductive source is considered in this study. Experimental data suggest that this is the dominant mode of coupling for most good quality shields. Figure 11 presents a section of multiconductor line with an overall shield. The transfer impedance of the shield is Z_{ab} and the external shield current is I_s such that the potential along the inside of the shield is

$$V_s = Z_{ab} I_s \Delta x . \quad (34)$$

It is assumed that the external current, I_s , is evenly distributed around the cable. The line integral of the electric field is related to the surface integral of the magnetic field by Stokes theorem:

$$\int E \cdot dl = -j\omega \int_{\Delta A} B \cdot dA , \quad (35)$$

where A is the area enclosed by the line integral. Assuming that the distances, Δx , and the wire-to-shield separation are small, the line integral of the electric field may be replaced by the potentials shown in figure 11:

$$V_i(x) - V_i(x + \Delta x) + V_s = -j\omega \int_{\Delta A} B \cdot dA , \quad (36)$$

where the field along the conductor is zero due to the assumption of perfectly conducting wires. Substituting equation (34) for V_s , dividing by x , and taking the limit as x approaches zero yields

$$-\frac{dV(x)}{dx} = -j\omega \int_0^h B \cdot dl - Z_{ab} I_s , \quad (37)$$

with h being the separation distance of the wire from the shield. The magnetic field, B , is the total field, incident and scattered. The scattered magnetic field may be related to the wire currents through self and mutual inductances. The incident magnetic field is the field present if the internal wires are not present. However, for frequencies of interest (≥ 100 MHz) and typical cable dimensions (radius < 2 cm) the cable without wires acts as a waveguide well below the cutoff frequency. Therefore, the incident magnetic field is zero, yielding the result that the voltage source on each wire is identical and equal to the product of the transfer impedance and the external current. This does not mean that the open circuit voltages at the ends of the cable are equal, simply that the per unit excitations are equal.

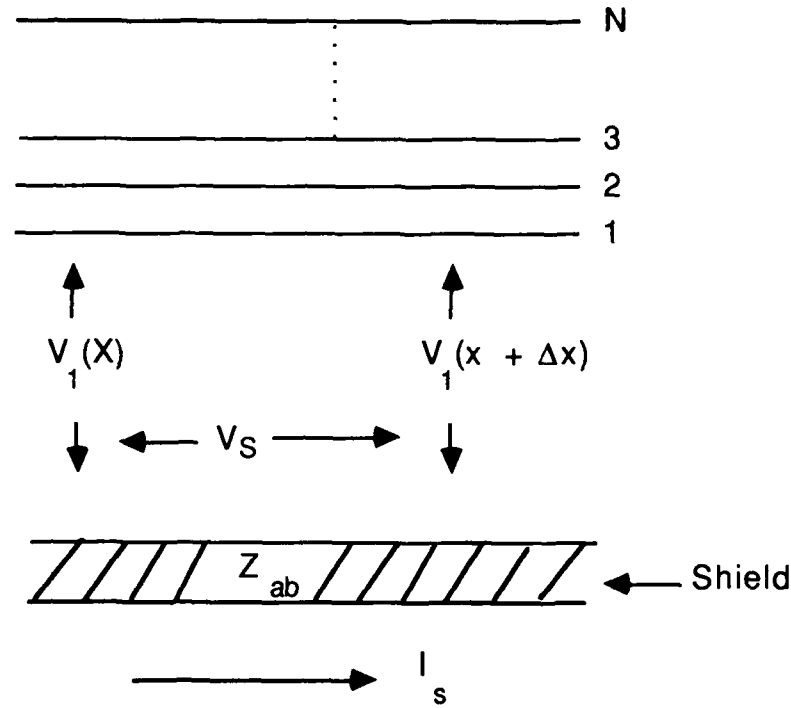


Figure 11. Multiconductor transmission-line segment.

Some insight may be obtained into the open circuit voltage response if a slightly different solution representation is used instead of equations (21), (22), and (24). The chain matrix formulation provided by Paul [10] is as follows:

$$\begin{bmatrix} \bar{V}(l) \\ \bar{I}(l) \end{bmatrix} = \bar{\Phi}(l) \begin{bmatrix} \bar{V}(0) \\ \bar{I}(0) \end{bmatrix} + \int_0^l \bar{\Phi}(l - \xi) \begin{bmatrix} \bar{V}_s(\xi) \\ \bar{I}_s(\xi) \end{bmatrix} d\xi = \bar{\Phi}(l) \begin{bmatrix} \bar{V}(0) \\ \bar{I}(0) \end{bmatrix} + \begin{bmatrix} \hat{V}_s(l) \\ \hat{I}_s(l) \end{bmatrix}, \quad (38)$$

where

$$\begin{aligned} \hat{V}_s(l) &= \int_0^l \left\{ \bar{\Phi}_{11}(l - \xi) \bar{V}_s(\xi) + \bar{\Phi}_{12}(l - \xi) \bar{I}_s(\xi) \right\} d\xi \\ \hat{I}_s(l) &= \int_0^l \left\{ \bar{\Phi}_{21}(l - \xi) \bar{V}_s(\xi) + \bar{\Phi}_{22}(l - \xi) \bar{I}_s(\xi) \right\} d\xi \end{aligned} \quad (39)$$

and

$$\begin{aligned}
\bar{\Phi}_{11}(l) &= \bar{C}^{-1} \bar{T} \cosh(j\omega\bar{\gamma}l) \bar{T}^{-1} \bar{C} & \bar{\Phi}_{21}(l) &= -\bar{T} \sinh(j\omega\bar{\gamma}l) \bar{T}^{-1} \bar{Z}_c^{-1} \\
\bar{\Phi}_{12}(l) &= -\bar{Z}_c \bar{T} \sinh(j\omega\bar{\gamma}l) \bar{T}^{-1} & \bar{\Phi}_{22}(l) &= \bar{T} \cosh(j\omega\bar{\gamma}l) \bar{T}^{-1}.
\end{aligned} \tag{40}$$

The open circuit voltage may be found from equations (36) through (38) if we simply let $I(0)$ and $I(l)$ equal zero.

Now consider the simple condition where the current source I_s is negligible, and also assume that the external shield current is constant over the length of the cable. Then the open circuit voltage at the $x = 0$ end becomes

$$\begin{aligned}
\bar{V}(0) &= \left[\bar{\Phi}_{21}^{-1}(l) \int_0^l \bar{\Phi}_{21}(l - \xi) d\xi \right] \bar{V}_s \\
&= \left[\bar{Z}_c \bar{T} \left[\sinh j\omega\bar{\gamma}l \right]^{-1} \int_0^l \sinh j\omega\bar{\gamma}(l - \xi) \bar{T}^{-1} \bar{Z}_c^{-1} d\xi \right] \bar{V}_s \\
&= \bar{Z}_c \bar{T} \left[j\omega\bar{\gamma} \sinh j\omega\bar{\gamma}l \right]^{-1} \left[\cosh j\omega\bar{\gamma}l - 1 \right] \bar{T}^{-1} \bar{Z}_c^{-1} \bar{V}_s.
\end{aligned} \tag{41}$$

The open circuit voltage at the other end is identical except for the sign. If the dielectric is homogeneous then the transformation matrix \bar{T} is the unit matrix and the eigenvalue matrix, $\bar{\gamma}$, becomes

$$\bar{\gamma} = \frac{1}{v} \bar{U}, \tag{42}$$

where v is the propagation velocity of the line. Then the open circuit voltage becomes

$$\bar{V}(0) = \frac{v \left(\cosh \frac{j\omega l}{v} - 1 \right)}{j\omega \sinh j \frac{\omega l}{v}} \bar{V}_s, \tag{43}$$

which means that the open circuit voltage is constant for all the wires if the source voltage, \bar{V}_s , is constant. The open circuit voltage at $x = l$ is identical except for the sign. A similar result may be obtained for the current-source-only condition, except the sign does not change from end to end.

If the cable is not homogeneous then there may be distinct eigenvalues. Since the eigenvalue matrix \tilde{Y} is diagonal, equation (41) may be rewritten as

$$V(o) = \bar{Z}_c \bar{T} \begin{bmatrix} \frac{v_1 \left(\cosh \frac{j\omega l}{v_1} - 1 \right)}{j\omega \sinh j \frac{\omega l}{v_1}} & & 0 \\ & \ddots & \\ 0 & & \frac{v_N \left(\cosh \frac{j\omega l}{v_N} - 1 \right)}{j\omega \sinh \frac{j\omega l}{v_N}} \end{bmatrix} \bar{T}^{-1} \bar{Z}_c^{-1} \bar{V}_s, \quad (44)$$

and the open circuit voltage vector will not be constant even if V_s is constant. The amplitude of each mode is proportional to its propagation velocity, v_i , and the wire potentials are weighted sums of these mode voltages with the weights determined by the T matrix and the characteristic impedance matrix, \bar{Z}_c .

Therefore, under the assumptions of inductive coupling dominance and a homogeneous dielectric the open circuit voltage from each wire to the shield will be equal. This means that a single source may be used in series with the cable shield at each end of the cable. This allows a significant reduction in the complexity of a Thevenin equivalent network for multiconductor cables. Now only two sources are needed, not $2N$. For nonhomogeneous dielectrics, the variation in wire to shield potentials may be small enough to allow use of only one source at each end. The suitability of this approximation will depend on the specific characteristics of the cable under study. Implementation details are discussed in the following section.

3.3 Equivalent Excitation Implementation

There are at least three implementation techniques that can be used in the simulation of the EMP response of a shielded cable. The purpose of discussing implementation techniques here is to provide theoretical insight into each method and to discuss the advantages and disadvantages of each. The discussion is limited to implementation on physical systems since circuit analysis modeling is relatively straightforward.

If it is known that the open circuit voltages are widely different for each of the N conductors, then the only implementation possible is the development of independent sources for each of the $2N$ ports. These sources must be reasonable approximations to ideal voltage sources. This requires the

source impedances to be much smaller than the characteristic impedances of the cable. The sources are placed in series with each conductor.

Of most interest here is the condition discussed in section 2.2.4, where the open circuit voltages are all approximately equal at each end of the cable. Then only two sources are needed, one for each end. These sources must be inserted into the reference conductor (the cable shield). The simplest way to implement these sources is to insert very low resistance in series with the cable shield and create the required voltage waveform across these resistors; see figure 12. The sources must provide currents defined by

$$I_s(x) = V_{oc}(x)/R, x = 0, 1, \quad (45)$$

where the resistance, R , must be much less than the bundle-to-shield impedance of the cable. The key advantages of this method are low power requirements and ease of specifying the source waveforms. The major disadvantage is that the shield continuity is disrupted by the insertion of the resistors.

A tradeoff may be made between the advantages and disadvantages of the simple resistive insertion technique. The integrity of the shield may be preserved if a segment of the shield is used instead of resistors to obtain the low source impedance. The source current waveform is then defined by

$$I_s(x) = V_{oc}(x)/(R_s + j\omega L_s)dx, x = 0, 1, \quad (46)$$

where R_s and L_s are the per unit length shield resistance and inductance, and dx is the length of the source segment. Equation (46) assumes that the shield excitation is dominated by the resistive-inductive transfer impedance and that the segment length, dx , is short compared to the smallest wavelength of the open circuit voltage. The preservation of the shield is obtained at the expense of the source current requirement. Typical shield transfer impedances range from 1 to 10 milliohms per meter, whereas the source resistance, R , may be as large as 1 ohm in some instances and still satisfy the requirement of being much less than the bundle-to-shield impedance. Therefore, the source current requirements may go up as much as an order of magnitude or more. The sources can be connected to the shield segments if some of the insulation at the ends of the cable is removed or if a short extension is added to the cable. The main advantage of this method is the preservation of the external shield integrity. The primary disadvantage is the high source current required, with a secondary problem being the difficulty associated with the physical connection to the shield.

The third technique attempts to reduce the disadvantages of the second technique through excitation of the entire cable shield as shown in figure 13. This technique has a much more complicated source specification because of the very distributed nature of the coupling through the shield. The source current is given in the frequency domain by

$$I_s(x) = V_{oc}(x)/V_d(x), x = 0, 1 \quad (47)$$

where $V_d(x)$ is defined as the open circuit bundle voltage due to an impulse function excitation (constant frequency spectrum). An analytical expression for $V_d(x)$ may be obtained from equation (4), with the voltage and current sources being functions of the external response to the impulse functions. This analysis is similar in complexity to the normal EMP analysis of the cable. It is also possible to determine $V_d(x)$ experimentally. The advantages offered by this technique are a slightly better source power requirement over the shield segment technique and complete maintenance of the shield integrity. The disadvantages are (1) a more complex source specification and (2) the possibility that the source current may not be a realizable function due to nulls or peaks in $V_d(x)$.

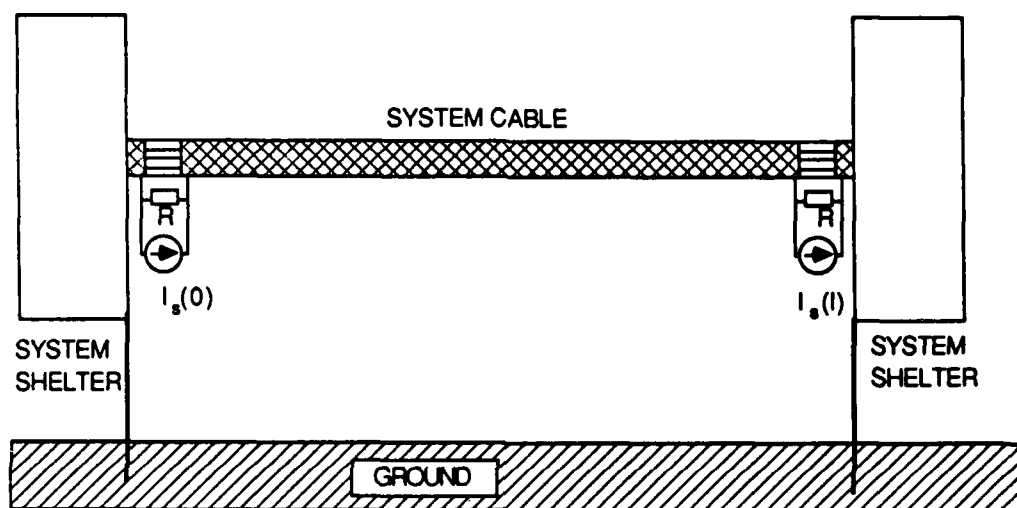


Figure 12. Equivalent source implementation using resistors in shield.

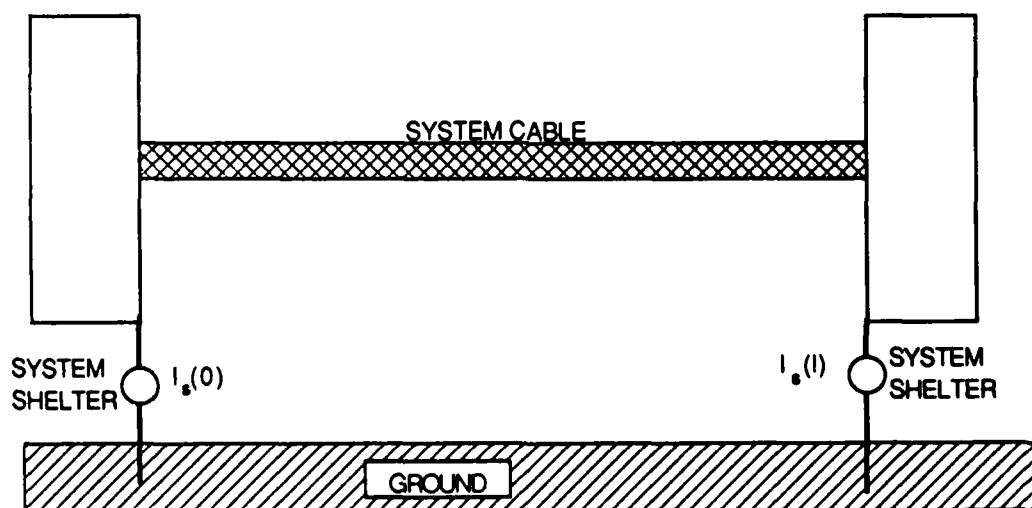


Figure 13. Equivalent source implementation over entire cable shield.

4. CONCLUSION

Efficient equivalent source representations were developed for coaxially shielded cables and for multiconductor cables. The correctness of these equivalent source representations was validated for linear and nonlinear terminations. The distribution of shield coupling parameters was studied for shielded multiconductor cables. The results of this study indicate that an approximate equivalent source excitation may be used for multiconductor cables, thereby reducing the source requirements even further. The approximate equivalent source excitation requires only two independent sources to represent the EMP response of the cable. The simplicity of the equivalent source excitation means that numerical models will run quicker, and system-level stimulation is possible with a lower power level and less complex simulators.

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